MARGINAL OSCILLATOR FOR A MODIFIED FREE PRECESSION MAGNETOMETER

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THESIS

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FOR A

MODIFIED FREE PRECESSION MAGNETOMETER

by

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June 1969

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by

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Submitted in partial fulfillment of the
requirements for the degree of

MASTER OF SCIENCE IN ELECTRICAL ENGINEERING

from the

NAVAL POSTGRADUATE SCHOOL
June 1969

ABSTRACT

The primary interest of the Naval Postgraduate School concerns the use of a magnetometer for mine detection, anti-submarine warfare, salvaging and other related naval operations.

The original concept of a modified free precession magnetometer utilyzing the Overhauser Effect was formulated by A. Abragam. Independent research on this magnetometer was conducted at the Naval Postgraduate School by P.E. Burcher and R.G. Landrum fulfilling requirements for their master's thesis in 1965. The objective of this thesis was to develop an improved marginal oscillator for the magnetometer.

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ACKNOWLEDGEMENTS

The author wishes to express his appreciation to his wife for her devotion and patience, to Dean Carl E. Menneken for his enthusiasm, advice and continuing encouragement, to Dr. William M. Tolles for his explanation of the physics and chemistry principles involved and to Hollis W. Oren for his assistance in circuit implementation.

I. INTRODUCTION

A. BACKGROUND

The application of the Overhauser Effect (dynamic polarization) to the earth's field nuclear resonance magnetometers requires the use of a marginal oscillator for signal detection. The marginal oscillator provides the sensitivity and narrow bandwidth necessary to extract the very weak nuclear signal. As will be seen in the preliminary theory, the marginal oscillator also provides phase coherence of the precessing protons in the sample nuclei.

The free precession nuclear resonance magnetometer has found extensive use in measuring the earth's magnetic field both for mineral prospecting and military applications. The idea of the free precession magnetometer was originally conceived by Dr. Russell Varian. He received a basic patent on the use of this technique in 1948. C.H. Bowen, Lieutenant, U.S. Navy, implemented a working system and devised the measurement instrumentation while on a three month industrial tour in 1954 with Varian Associates Research Laboratory at Palo Alto, California. He later received a master's degree at the Naval Postgraduate School with his work on the free precession magnetometer fulfilling the requirements for his thesis [3]. A. Abragam in 1957 suggested that by use of the Overhauser effect continuous output information could be obtained [1]. Because of the inherent low data rate of the free precession magnetometer, research was conducted in a later thesis [2] on a modified free precession magnetometer utilyzing the Overhauser Effect to provide a continuous data rate for increased measurement accuracy. This modified magnetometer required the addition of a marginal oscillator for signal detection which is the subject of this thesis.

B. PRELIMINARY THEORY

Detailed theory concerning the free precession and modified free precession magnetometers can be found in the previously mentioned references. However, a brief discussion of the basic theory involved is reiterated in this thesis to provide a clearer understanding of the signal detection sub-system.

Nuclear magnetic resonance exists due to the fact that atomic nuclei possess angular momentum and magnetic moment. This phenomenon is depicted in figure 1. Every proton possesses gyroscopic characteristics in that they have a magnetic moment $\bar{\mu}$ and angular momentum $\bar{A}$, which are collinear. The sum of the individual magnetic moments yields a total magnetic moment $\bar{M}$ which is given by the product of $\bar{\mu}$ and the number of nuclei in the sample material. This total magnetic moment vector will precess about an applied field vector $\bar{H}_0$ at an angular precession frequency known as the Lamor frequency. The precession frequency is computed as follows:

$$f_0 = \frac{\gamma}{2\pi} H_0 \qquad (1)$$

where γ is the "gyromagnetic ratio" (charge/mass) for protons $\frac{\gamma}{2\pi}$ = 4.2576 khz/gauss. Therefore, $f_0 = 4.26 \ H_0$

The free precession magnetometer consists primarily of a detection coil surrounding a sample of material containing a large number of protons. This is depicted in block diagram form in figure 2. A pulsed dc current, sufficient in magnitude to create a magnetic field much greater than the earth's field, is injected into the coil. This field produces polarization of the proton spins in the atomic hyperfine structure. During the time that the polarization pulse is applied, the resultant magnetic moment $\bar{M}$ of the sample nuclei will precess about the applied field $\bar{H}_0$ until it gradually comes into allignment with the field. When the pulse is removed, $\bar{M}$ will precess about the earth's field at a frequency directly proportional to the magnitude of the earth's field. The magnitude can be determined as in equation (1) with H_0 representing the earth's field. The frequency of precession can be determined electronically since the precession motion induces a voltage of a few micro-volts in magnitude in the coil. This voltage is amplified and the frequency measured by the use of a frequency counter. When the induced signal is viewed on an oscilloscope, it has the appearance of an exponentially damped sinusoid. The damping action is due to the magnetic moment vector decreasing exponentially with time.

The accuracy of the earth's field measurement depends largely on the accuracy with which the signal frequency is measured. If the measurement time is increased the accuracy of the measurement correspondingly increases. The measurement time allowed is limited by the rate of signal decay with time. If the signal could be maintained at a constant level between the applied dc pulses, and not be damped, the measurement accuracy would be increased.

Electrons exhibit similar resonance characteristics as the protons except, due to their much lower mass, the precession frequency is higher. It can be seen in equation (1) that the precession frequency is inversely proportional to the particle mass. In general, the precession frequency for electrons is in the MHz range and that for the protons is in the KHz range. It should also be mentioned that electron resonance properties exhibit a very strong electrostatic interaction with the lattice, while the protons in the nucleus are in thermal equilibrium with the lattice. The protons are shielded from the lattice by the electrons. It is this property that becomes extremely important in the understanding of the Overhauser Effect.

Basically, the Overhauser Effect states that electronic energy transitions cause nuclear energy trnasitions and dynamic polarization of the proton spins is thus obtained. The proton precession signal is enhanced by induced electronic transitions. RF magnetic energy of sufficient strength and at a given pre-determined frequency will be absorbed by electrons in certain energy states. The source of the energy is referred to as the "pump" and the pumping frequency required to obtain the desired effect can be calculated. The electron-nucleus coupling constant can be obtained by means of an electron spin resonance (ESR) experiment, and the pumping frequency required can be calculated using a reference sample (Fermi-salt) for which a precise coupling constant and pumping frequency are known. The following ratio produces the unknown frequency:

$$\frac{f_p\ (\mathrm{MHz})}{A} = \frac{54.731}{13.6}$$

where "A" is the coupling constant. The energy absorbed by the electrons raises them to a higher energy state, thus causing an electron transition. Proton energy transitions will occur simultaneously, as explained previously, and the sample nuclei are completely shielded from the applied field. It can be seen that since the nuclei are shielded from the applied field inducing polarization, precession of the protons about the earth's field vector can occur continuously. In a nuclear resonance experiment the observed signal is proportional to the energy level population difference, therefore the application of the RF energy tends to greatly enhance the nuclear resonance signal. This enhancement due to dynamic polarization is referred to as the Overhauser Effect.

The modified free precession magnetometer described above is depicted in block diagram form in figure 3. The precession signal to be detected is extremely small in amplitude and is obscured by the existing noise level. Therefore, it is important to couple the maximum signal to the electronic detection system and at the same time limit the bandwidth for noise suppression. It was found that a marginal oscillator, operating in a marginal condition, effected a high "Q" detection coil, limiting the frequency pass band and increasing signal amplitude. The center frequency (f_0) of the marginal oscillator input tank circuit must correspond very closely to the precession signal frequency for optimum performance. While in a marginal state, the marginal oscillator produces phase coherence of the precessing protons by feeding back a small amount of energy at the precession frequency. Thus, it synchronizes the precession of a large number of protons in the sample nuclei. The phase coherence reduces signal cancellation due to random motion of the precessing protons. A buffer amplifier and bandpass filter are introduced at the marginal oscillator output to reduce loading effects and further increase the signal-to-noise ratio respectfully.

In summary, three basic requirements are placed on the marginal oscillator:

1. Produce phase coherence for the precession signal
2. Amplify the signal
3. Suppress noise by narrowing the frequency passband

II. MARGINAL OSCILLATOR

A. GENERAL

The marginal oscillator is an amplifier with positive feedback applied between the output and input of the amplifier. It is depicted in block diagram form in figure 4. The feedback gain is made variable for adjusting the oscillator to the marginal condition. The marginal condition is realized when the feedback voltage is increased to the critical voltage as can be seen in figure 5. If the feedback voltage increases beyond V_C, the amplifier is in self-sustained oscillation. This condition is undesirable since the marginal oscillator is no longer an amplifier sensitive to the input signal.

It can be seen in figure 4 that a parallel L-C tank circuit is placed between the input signal and the marginal oscillator input. The quality factor (Q) of the tank circuit is defined as the energy stored divided by the energy lost in the circuit. Due to the positive feedback in the marginal oscillator, the energy lost in the tank circuit is reduced and the Q is said to be synthetically increased. Therefore, if the energy lost is minimized, the Q is maximized allowing for maximum input signal amplification. This is true since the output voltage from a parallel tank circuit is Q times the input voltage. Of course, if the positive feedback is increased beyond V_C, the Q becomes infinite and the marginal oscillator is in self-sustained oscillation. This can be seen in figure 6. There is an optimum region of performance where the marginal oscillator acts as an amplifier and allows maximum voltage gain of the input signal. When the marginal oscillator is in this region it is in the marginal condition previously referred to.

If the frequency of the input signal corresponds exactly to the center frequency f_o of the tank circuit, maximum amplification is realized. However, in many cases, the input signal is slightly changing in frequency. As the signal frequency deviates from f_o the signal amplification decreases until the signal is no longer in the frequency passband Δf. It is for this reason that the increased Q corresponding to a decreased bandwidth must allow for sufficient signal frequency deviation without a great loss in gain. The frequency deviation allowed for and the Q desired are dependent upon the specific application and the design criteria.

B. REQUIREMENTS

The marginal oscillator used previous to this thesis utilized vacuum tubes which required a high plate voltage and filament heating voltage. Additionally, the marginal oscillator required operator manipulation during the data taking process. Since the measurement time can be quite lengthy, an automatic operation to maintain optimum performance is required for a practical system. A compact system was needed since many separate units requiring individual power supplies produced a relatively cumbersome magnetometer. In the eventuality that this magnetometer is used in an aircraft, weight, size and power requirements become paramount considerations.

Since the magnetometer must be able to respond to small changes in the earth's magnetic field, the marginal oscillator must have the ability to track a fractional change in the precession signal frequency. In view of the exceedingly low signal voltage, the marginal oscillator must be inherently a low noise device. The control of the marginal state of the oscillator must be fast and precise in order to maintain the desired accuracy continually. The above mentioned items were taken into consideration in the selection of an optimum marginal oscillator.

C. SELECTION

It was decided that the active device chosen be solid-state in nature since solid-state devices are small in size and require a minimal power supply. Also, the solid-state device provides longer life time since no filaments are required and the ability to withstand greater pressure and shock. Of course, the selection of this type of device for the active element in the circuit introduces a definite need for temperature considerations since these devices are temperature sensitive and wide temperature variations must be expected.

The first marginal oscillator considered is shown in figure 7. This is a simple "twin T" oscillator with the positive feedback gain varied by R3. The circuit could also be utilized as a bandpass filter with slight modifications. It did function as a marginal oscillator but had certain disadvantages. The input impedance to the device was not sufficiently large and loaded the input tank circuit. As a result the Q of the circuit was reduced considerably and the gain of the input signal was not sufficient. It was also very difficult to adjust the oscillator to the marginal condition. It was then considered to utilize the twin T circuit as a bandpass filter, but the bandpass could not be made sufficiently narrow. A bandpass of ± 30 Hz either side of the center frequency was desired.

Integrated circuits were then investigated as the active device in the marginal oscillator. Also, a circuit with a high degree of stability inherent in the design was sought in an attempt to eliminate the need for an automatic feedback control circuit. Thus, the circuit of figure 8, utilizing an operational amplifier, was designed. The negative feedback, applied to the inverting input of the amplifier through R3, stabilizes the gain and makes it essentially independent of the integrated circuit's characteristics. At the same time, the RC network formed by R1, C1, R2, and C2 applies positive feedback to the non-inverting input. If the positive feedback is equal to or greater than the negative, the circuit will oscillate at the frequency at which the phase shift through the RC network is zero. If R1 x C1 = R2 x C2 the frequency of oscillation is 1/2 R1C1. Thus, with sufficient negative feedback, it is possible to obtain a circuit whose characteristics do not change with age or temperature, but only with the precision of the external elements.

The circuit provided the marginal oscillator characteristics but was found to be too difficult to adjust. Additionally, precision components were not available to allow a rigid test of stability. It was then decided to search for another integrated circuit device which would enable the use of a feedback control element without loss of sensitivity.

Integrated circuit differential amplifiers were next investigated. Certain integrated chips have temperature compensating networks inherent in their construction such that the circuit will operate over a wide range of temperatures before its characteristics are degraded. Also, on some integrated chips a constant current generator is incorporated to supply the current to the emitter-coupled pair of transistors. Since the transconductance of the amplifier is directly proportional to the current supplied, a control circuit can be readily employed to vary the current and thus the amplifier gain. If the gain increases, the device will break into self-sustained oscillation. The control element will then sense the increase in output and reduce the emitter current which in turn reduces the gain of the amplifier to the desired level.

The R.C.A. CA3028 and CA3000 integrated circuits were selected since the constant current generator with external tie points for automatic gain control was inherent in the chip. The circuit diagrams for both integrated circuits are shown in figure 9. The CA3000 was ultimately chosen since it possessed internal temperature compensation via D1 and D2 in the constant current generator and the emitter resistors R4 and R5 increased the stability.

C. DESIGN

1. Differential Amplifier Theory

The RCA 3000 integrated differential amplifier circuit was used as the active device of the marginal oscillator. As mentioned previously, the circuit has a controlled constant current source. Additionally, a temperature compensating network is incorporated as an integral part of the controlled-source circuit to assure that the circuit gain, dc operating point, and other important characteristics vary as required over the operating temperature range.

As indicated in figure 10, the differential amplifier is usually operated from dual dc supply voltages, + Vcc and - Vee, applied in the normal polarities used for n-p-n transistors.

Single-supply operation is also feasible, but requires an external voltage divider network and an additional bypass element. Single-ended outputs may be coupled from the collector of either transistor of the differential pair. In the design, one output was used for the positive feedback network and the other was used to provide the oscillator output. In this way the feedback network was not loaded down by additional stages.

In the differential amplifier, the sum of the emitter currents of the differential pair of transistors, Q1 and Q2, must be equal to the total amount of current (I_o) supplied to the constant-current sink. This can be seen in figure 7. As shown by the application of Kirchoff's second law at node e:

$$I_{e1} + I_{e2} = I_o$$

The emitter - to - base voltages V_{be1} and V_{be2} of the emitter-coupled transistors are expressed as follows:

$$V_{be1} = V_{b1} - V_e \tag{3}$$

$$V_{be2} = V_{b2} - V_e \tag{4}$$

where V_e is the voltage at node e and V_{b1} and V_{b2} are defined as indicated in figure 10. If the V_e term is eliminated in equations (3) and (4), the following result is obtained:

$$V_{b1} - V_{b2} = V_{be1} - V_{be2}$$

This latter equation defines the differential input voltage for the differential amplifier. With the base of Q_2 grounded, single-ended operation is obtained. The outputs from the collectors of the differential pair of transistors are identical in magnitude but 180 electrical degrees out of phase. The collector currents are given as follows:

$$I_{c1} = \frac{\alpha \; I_o}{1 + \text{Exp}(-) \; V_{b1}/h}$$

$$I_{c2} = \frac{\alpha \; I_o}{1 + \text{Exp}(+) \; V_{b1}/h} \qquad (5)$$

where $h = KT/q$

and K = Boltzmann constant

T = temperature in degrees Kelvin

q = charge on an electron

The transconductance evaluated at the operating point ($e_b = 0$) is given as follows:

$$g_m = q\alpha I_o/4KT \qquad (6)$$

It can be seen that g_m (the slope of the transfer curve) can be changed, without changing the linear input region, by varying the value of I_o. This relationship implies that automatic gain control is inherent in the differential amplifier when the current I_o is controlled. The emitter degeneration provided by the emitter resistors in the CA 3000 chip effectively reduces the gain (transconductance) of the differential pair of transistors, but also increases the linearity of the transconductance.

2. Oscillator Design Computations

Reference is made to figure 9(b) for the design. A dc path to ground was provided at terminals 1 and 6 through a 1K ohm resistor. This was made necessary for correct dc biasing of the amplifier. The positive feedback loop was provided between terminals

1 and 8. Terminal 8 was chosen as the output to provide the feedback since no phase reversal of the input signal exists between the input and this terminal. The supply voltages were chosen to be ± 6 volts for linear operation of the device. The AGC voltage (with negative polarity) is applied directly to terminal 2.

The positive feedback loop consists of a capacitor-resistor combination. The resistor is made variable to provide the variable feedback gain necessary to obtain a marginal condition without self-sustained oscillation. The following values were chosen for the components:

C = .01 micro-farad

R = 10K ohm (potentiometer)

The input tank circuit consists of the detection coil and a parallel capacitor. The inductance of the coil is 23 millihenry and a capacitor of 0.2 micro-farad was chosen to correspond to a resultant frequency of operation of 1.5 KHz. When the circuit is used with the magnetometer, capacitive decade boxes will be used to increase the frequency of operation to correspond to the local earth's field (approximately 2.2 KHz).

3. Automatic Feedback Control Circuit

The type of control necessary was first considered. Current sensing and rate-of-change of voltage were explored and found not to give the range and sensitivity of control necessary. The change of voltage from that caused by the input signal to that due to self-sustained oscillation was not sufficiently fast to produce a control signal. A voltage level comparator was decided upon and found to be a satisfactory method of controlling the marginal oscillator.

The voltage level at the output of the marginal oscillator due to an input signal is about 2 volts less than that due to self-sustained oscillation. An integrated circuit operational amplifier with a differential input is used as the comparator. A variable dc reference voltage is applied to the non-inverting input and the marginal oscillator output is rectified and fed to the inverting input of the operational amplifier. When the dc level produced by the marginal oscillator output signal exceeds that of the reference voltage, an output from the comparator will result. The comparator output is then connected to the AGC control of the oscillator. This operation is depicted in figure 11. The comparator output is clamped to ground (zero output) when the reference voltage level exceeds that of the dc signal level.

Buffering was necessary between the oscillator and the control circuit, therefore an integrated circuit operational amplifier utilizing a μA 741 was used as a buffer amplifier. The amplifier is depicted in figure 12. The input to the amplifier is taken from the second output of the marginal oscillator at terminal 10. The output is then rectified and connected to the inverting input of the voltage comparator.

The time constant of the RC filter used with the rectifying diode, as shown in figure 11, must be long compared to the ac signal. The period of the ac signal is about 0.5 milli-seconds (corresponding to 2 KHz), therefore the RC filter was designed to have a time constant at least ten times as great. The values used are as follows:

$C = 0.22$ micro-farad

$R = 30K$ ohm

The rectifying diode is a GE 1694.

The gain of the comparator is set by the negative feedback resistor R1 and is made variable to obtain the output voltage level necessary to control the marginal oscillator. Also, since the oscillator output signal level will change with time, the reference voltage must be made variable to correspond to this change. The reference voltage level is adjusted to be 0.2 volts greater in magnitude than the rectified signal level. Thus, if the marginal oscillator output increases in amplitude due to a noise spike or other cause, the voltage comparator will supply an AGC voltage to counter the voltage change as soon as the level changes by more than 0.2 volts. As can be seen in figure 8, D2 is forward biased when the positive reference voltage at the non-inverting input exceeds the signal voltage; thus, the output is essentially clamped to ground as mentioned earlier. Actually, however, the output is slightly above ground potential due to the small forward diode resistance (approximately 20 ohms). This offset in output voltage was not sufficiently great to affect the oscillator control and was neglected.

III. RESULTS

The signal seen at the marginal oscillator input is the proton precession signal induced in the coil multiplied by the Q of the tank circuit. Therefore, it is important to obtain the highest Q possible for maximum signal amplification. The Q was measured both with and without the marginal oscillator in the circuit. The merit of the marginal oscillator for signal amplification was thus realized. The values found were as follows:

without marginal oscillator: $Q = f_o/\Delta f$

$= 2.33 \text{ KHz}/1.09 \text{ KHz}$

$= 2.3$

with marginal oscillator in a marginal condition: $Q = f_o/\Delta f$

$= 2.48 \text{ KHz}/0.14 \text{ KHz}$

$= 17.7$

The frequency spread (Δf) was determined by noting the output signal voltage and the frequency at which it was reduced to 0.707 of its maximum value.

As mentioned previously, the signal voltage level at the marginal oscillator output due to input signal excitation is lower in magnitude than the level produced from self-sustained oscillation. This is the fact that enabled voltage level comparison to be used for the automatic gain control. For a given input voltage (micro-volts), the output voltage levels were measured as follows:

input signal excitation
with marginal oscillator
in marginal condition: V_{out} = 1.44 volts (RMS)

marginal oscillator in
self-sustained
oscillation: V_{out} = 2.30 volts (RMS)

For the input signal used in this experiment, the comparator reference voltage level was set as follows:

$$V_{ref} = 1.44 + 0.2 \text{ volts}$$
$$= 1.64 \text{ volts (RMS)}$$

This comparator reference voltage assures that the AGC voltage at the comparator output is zero when the marginal oscillator is not in self-sustained oscillation. Also, the output signal voltage is allowed to increase by 0.2 volts before the AGC control of the marginal oscillator commences.

The relative gain of the marginal oscillator was measured as a function of applied dc control voltage at terminal 2 in the constant current generator. The gain was measured in decibels (DB) with the gain corresponding to zero control volts normalized to zero DB. The resulting curve is shown in figure 13. It can be readily seen from figure 13 that a control voltage (comparator output voltage) of -4 to -5 volts is necessary to reduce the increased marginal oscillator gain while in self-sustained oscillation. The comparator output voltage was thus set at -4.5 volts to correspond to the maximum amplitude of the self-sustained oscillation.

An earth's field contour map indicated that a frequency variation from 1.5 KHz to 2.8 KHz might be expected throughout the world. It was desired to determine the change in marginal oscillator gain within this range of frequencies. Capacitive decade boxes were placed in parallel with the input tank circuit to raise the frequency of operation in increments of 0.5 KHz from 1.5 KHz. For each frequency change the marginal oscillator was adjusted to the marginal condition and the output voltage was measured. The gain was found to increase by as much as 3 DB. This change of gain is of no consequence since the reference voltage in the voltage comparator is variable and is made to compensate for the output signal amplitude increase. The desired effect is to have zero AGC voltage for the normal level of output signal voltage.

IV. CONCLUSIONS

As mentioned previously, the input signal used in the testing of the marginal oscillator was generated with a laboratory audio signal generator. It was inductively coupled (lightly) to the detection coil to simulate the precession signal which is in the order of micro-volts. Time did not allow obtaining the actual signal for the marginal oscillator evaluation. However, the characteristics of the oscillator proved to be superior to those previously observed in the vacuum tube version.

The marginal oscillator will be placed on a printed circuit board with the associated power supply. There will be adequate space available on the board to incorporate later the bandpass filter and other portions of the signal detection system if an integrated circuit design is used. Additionally, all of the circuits could utilize the same power supply. Ultimately it would be desirable to have the entire detection system in one small enclosure for compactness.

It was mentioned earlier that the marginal oscillator input tank circuit center frequency must be made to correspond exactly to the precession signal frequency. Optimum signal amplification is obtained for this condition. The precession signal frequency varies as much as 1.3 KHz depending on the geographical location of the magnetometer. An automatic tuning circuit which would keep the input tank circuit tuned very closely to the input signal, over this frequency range, is highly desirable since it would eliminate the need for manual tuning by the operator.

APPENDIX A . FIGURES

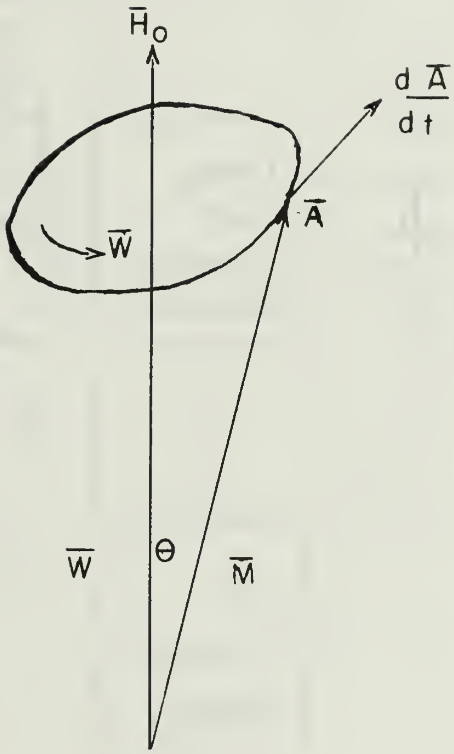


FIGURE 1

MAGNETIC VECTOR DIAGRAM

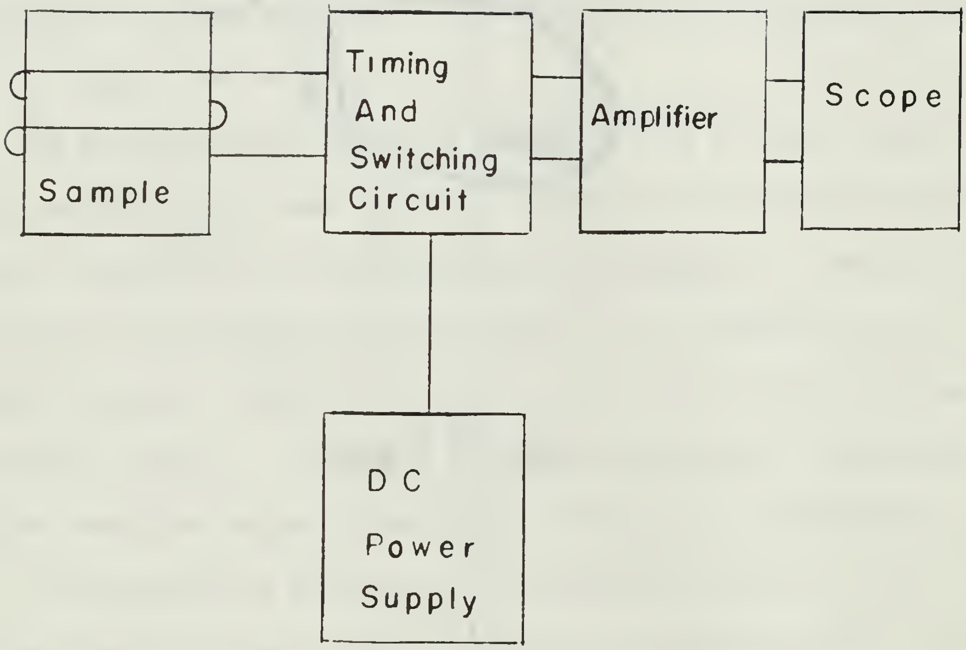


FIGURE 2

FREE PRECESSION MAGNETOMETER
BLOCK DIAGRAM

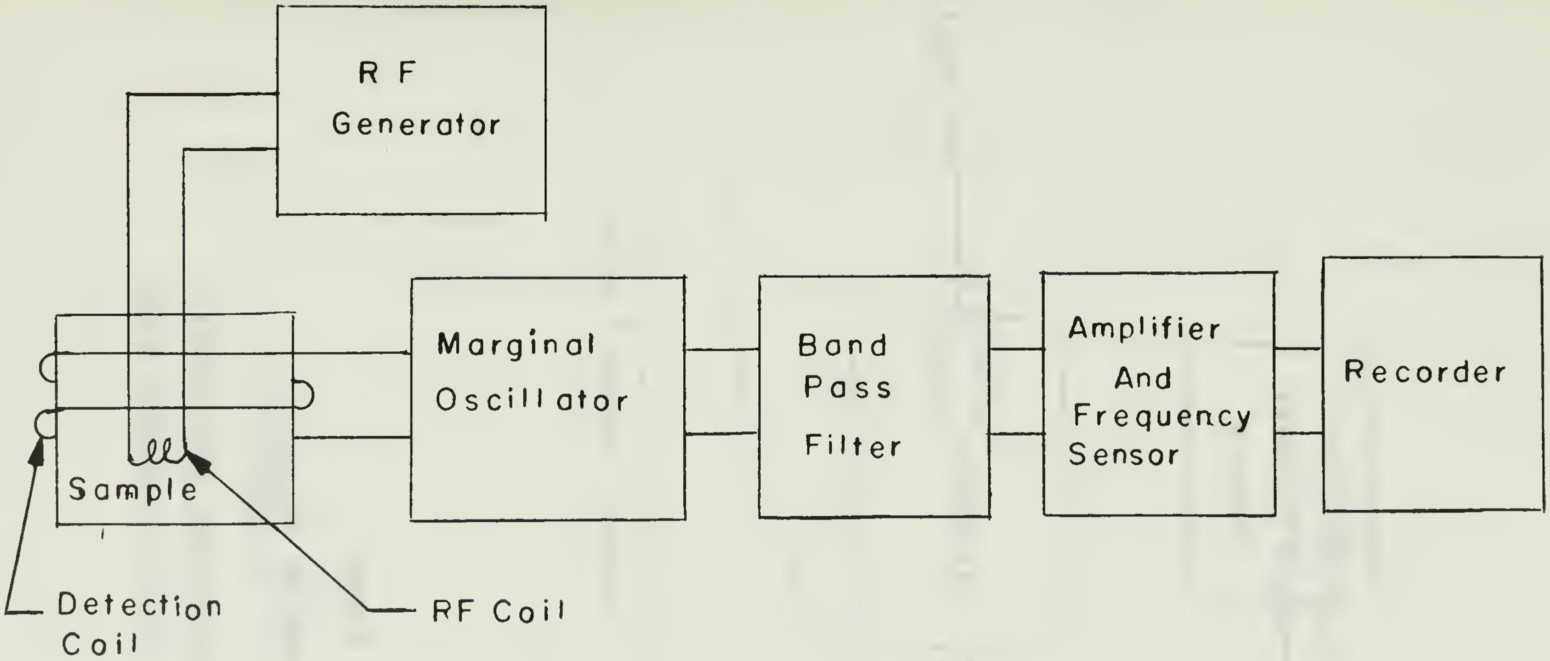


FIGURE 3

MODIFIED FREE PRECESSION MAGNETOMETER

BLOCK DIAGRAM

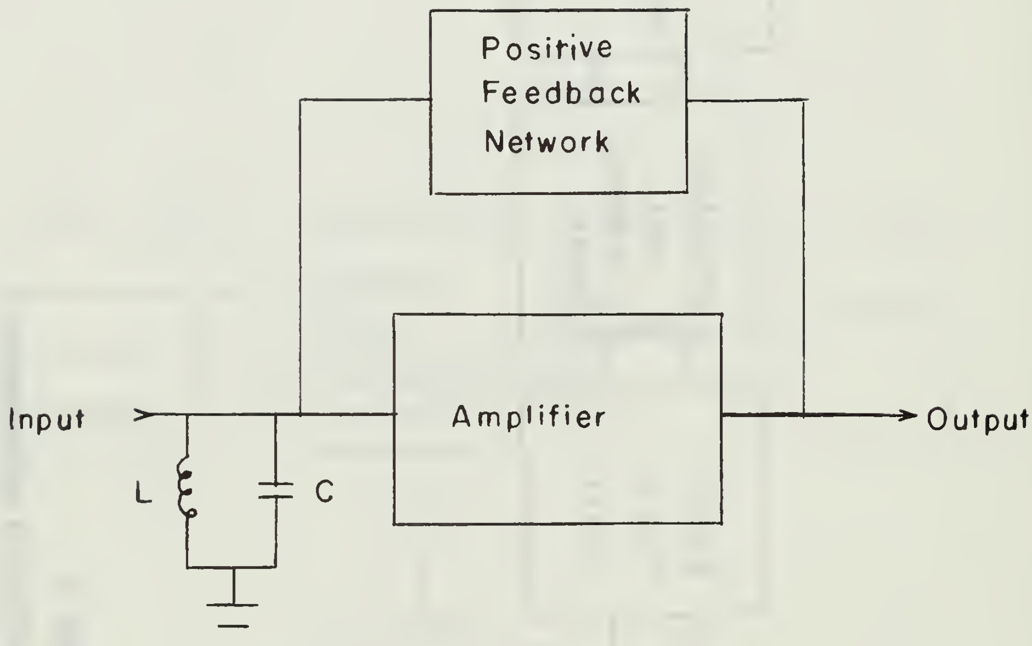


FIGURE 4

BASIC MARGINAL OSCILLATOR

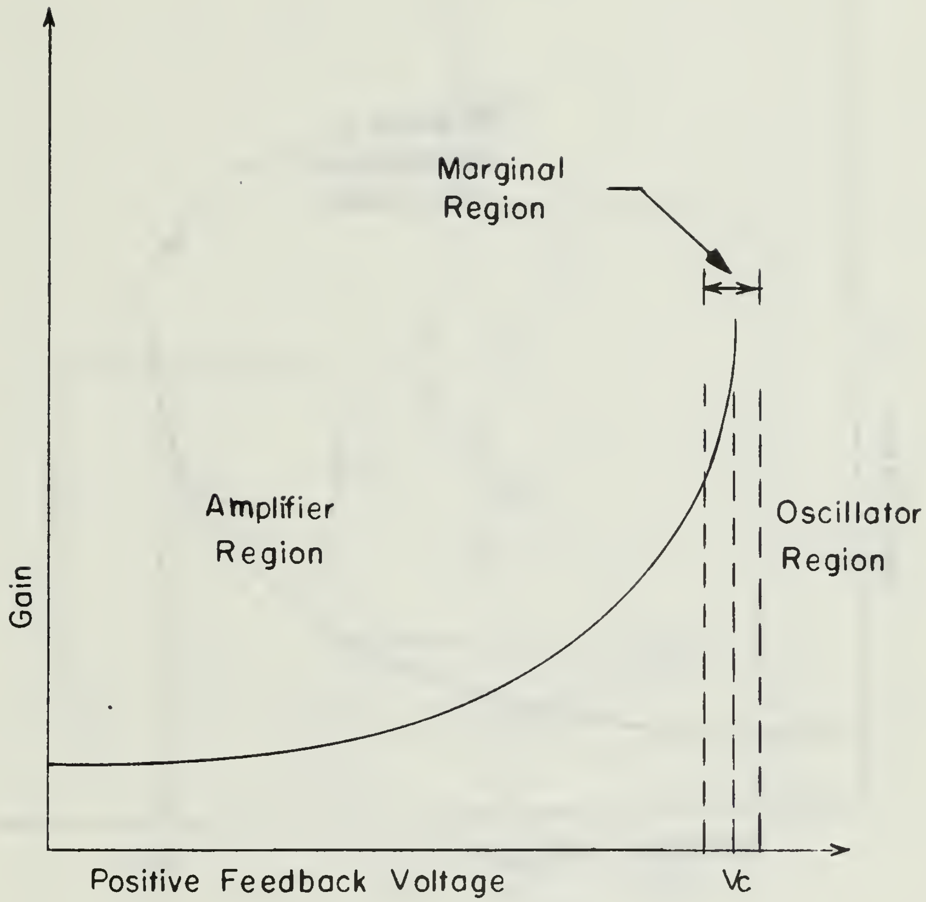


FIGURE 5

TYPICAL GAIN CHARACTERISTICS
FOR A MARGINAL OSCILLATOR

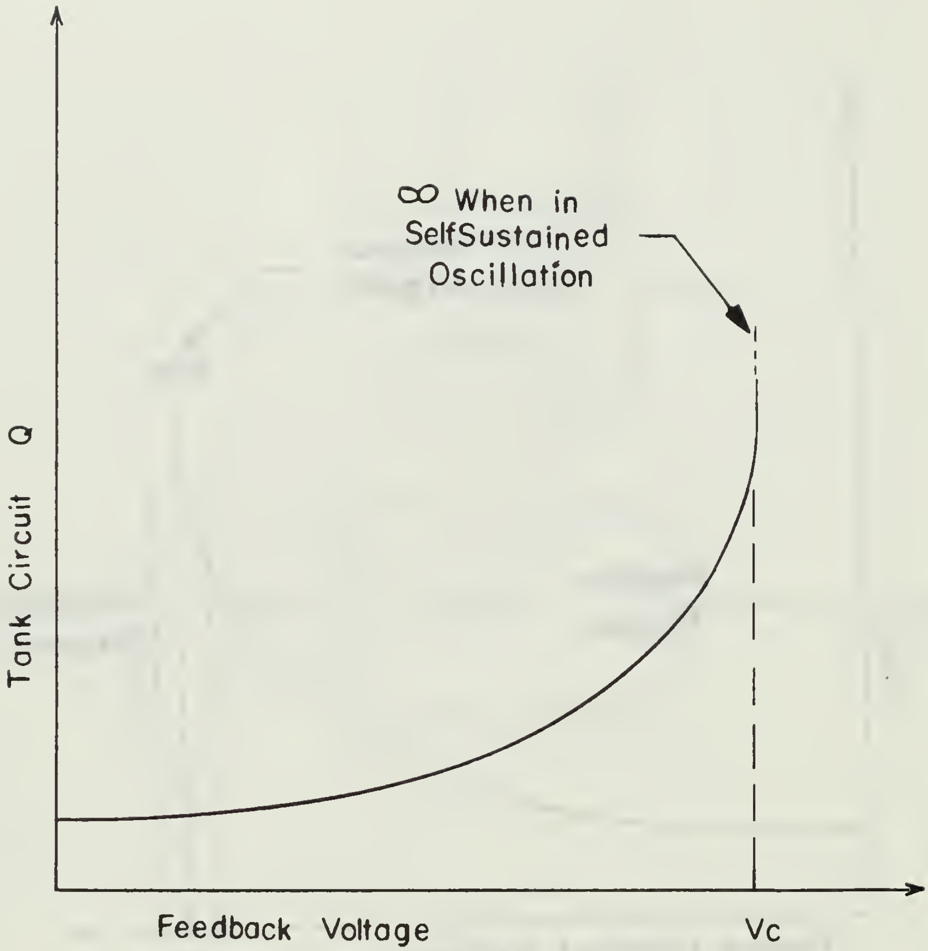


FIGURE 6

Q VARIATION WITH FEEDBACK VOLTAGE

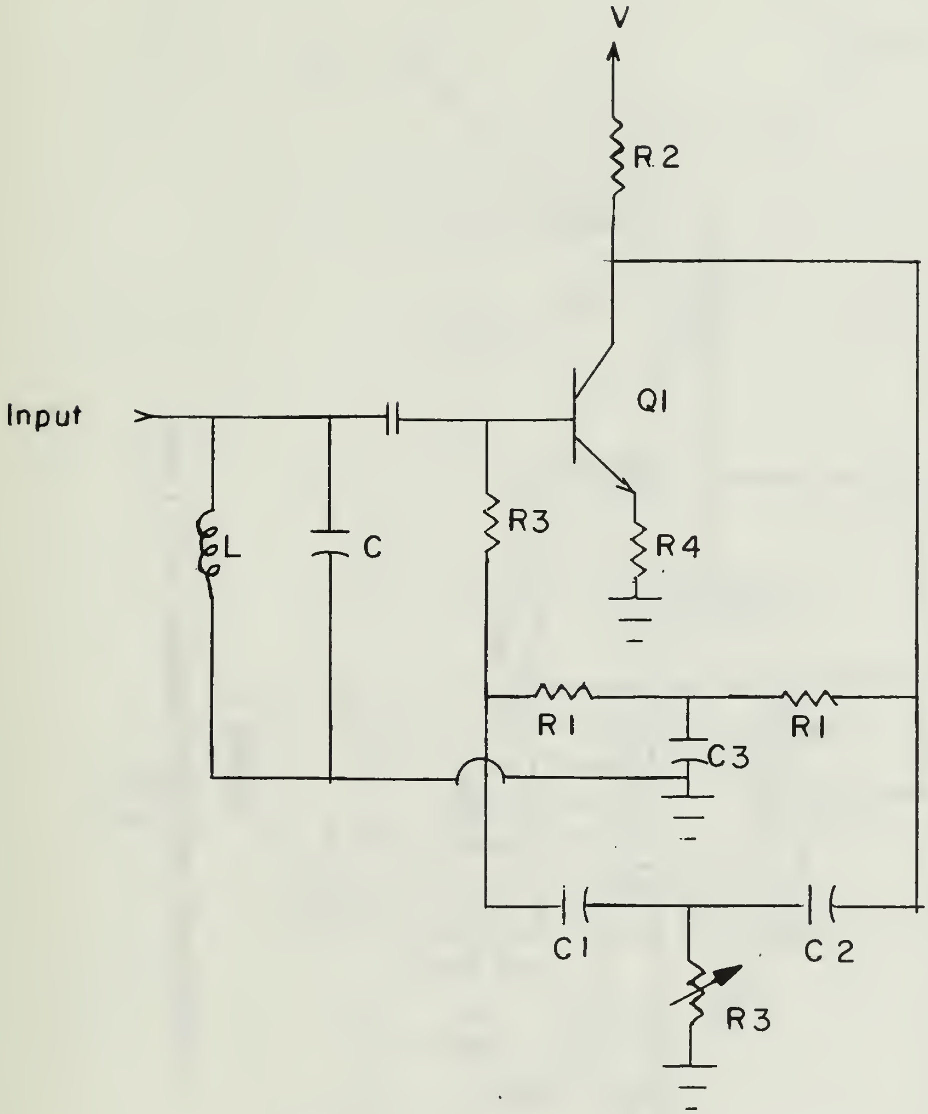


FIGURE 7

TWIN T MARGINAL OSCILLATOR

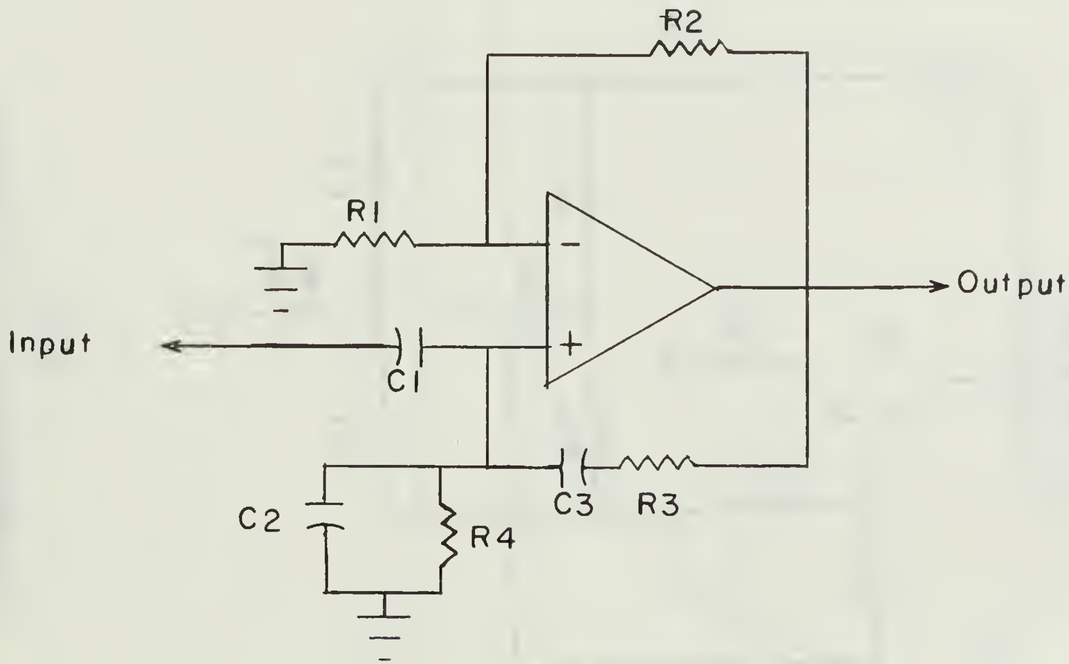
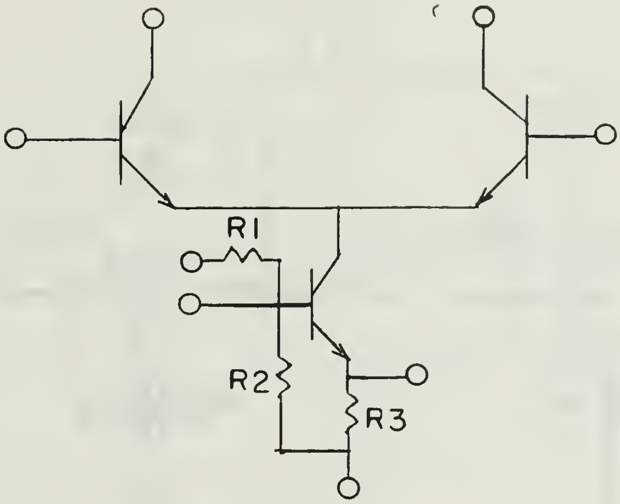
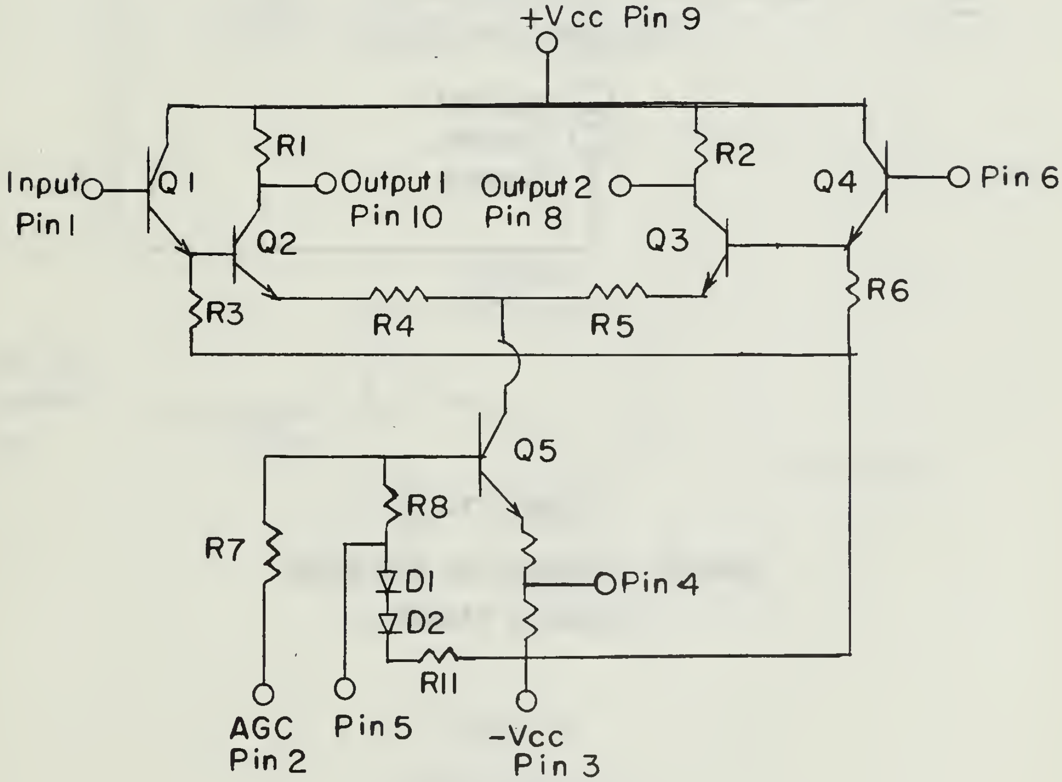


FIGURE 8

OPERATIONAL AMPLIFIER MARGINAL OSCILLATOR



CA 3000
(a)



CA 3000
(b)

FIGURE 9
I.C. DIFFERENTIAL AMPLIFIERS

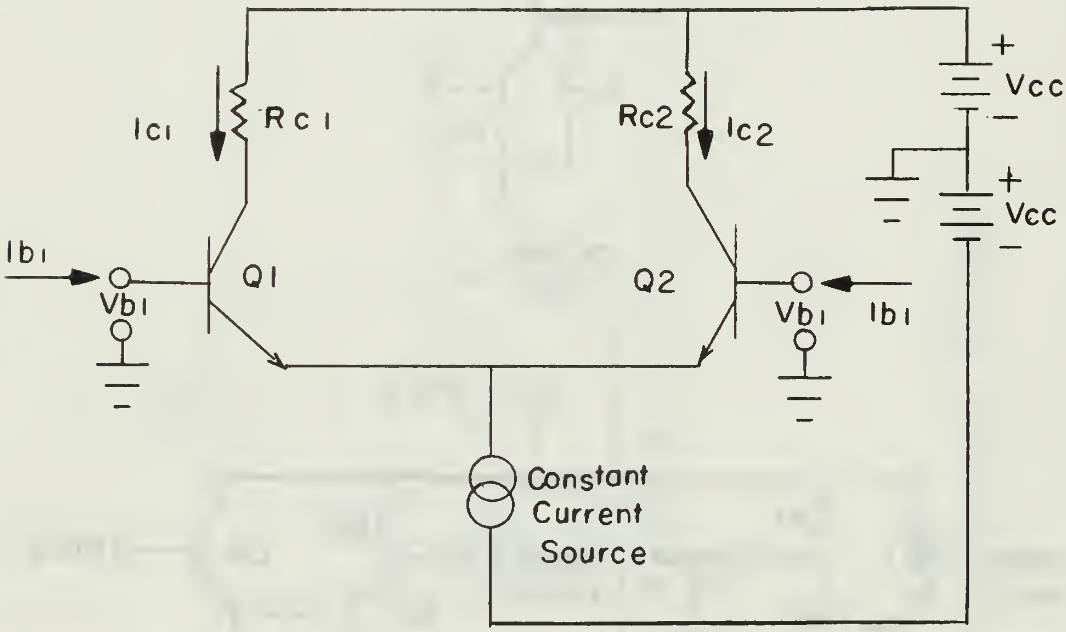


FIGURE 10

GENERAL DIFFERENTIAL AMPLIFIER
SCHEMATIC DIAGRAM

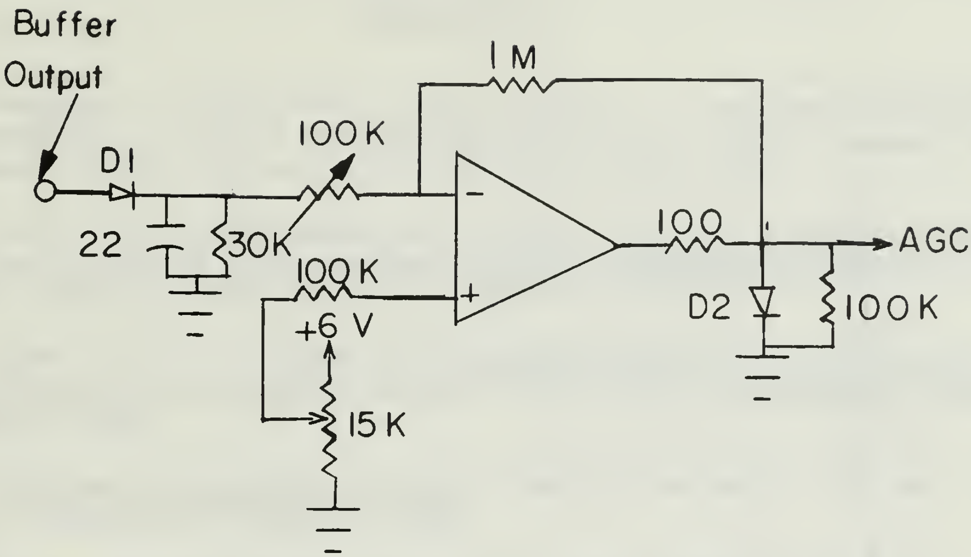


FIGURE 11

DC VOLTAGE COMPARATOR

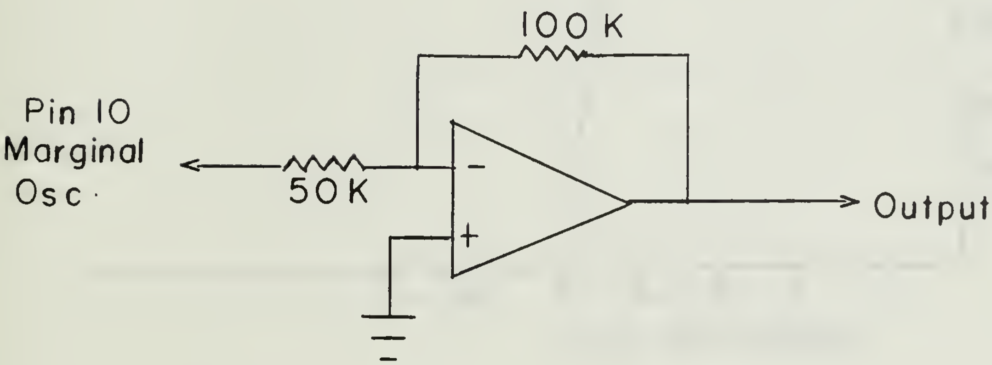


FIGURE 12

BUFFER AMPLIFIER

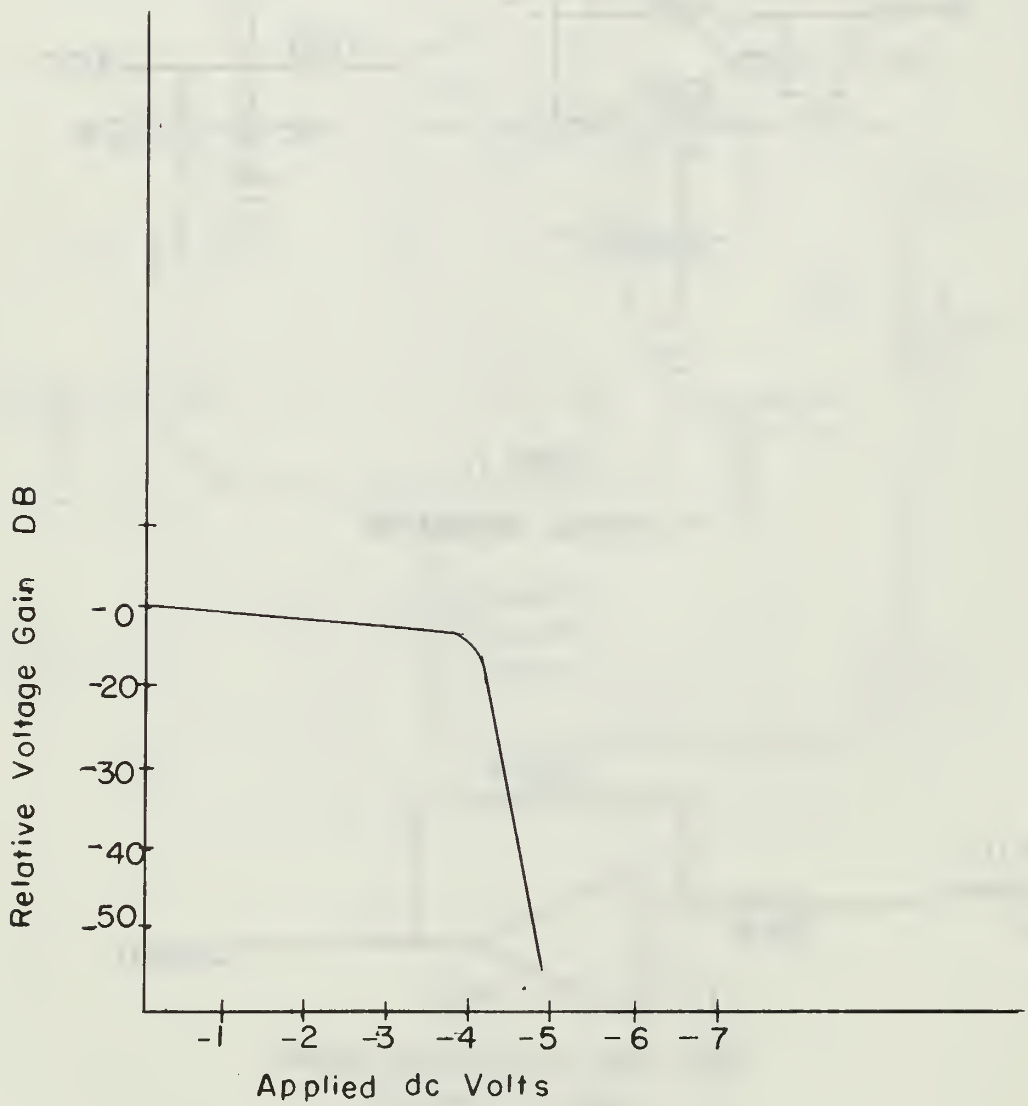


FIGURE 13

AGC CONTROL GRAPH

BIBLIOGRAPHY

1. Abragam, A., Combrisson, J. and Solomon, I., "Nuclear Polarization By Overhauser Effect in Solutions of Paramagnetic Ions", Academie Des Sciences, C.R.V. 245: no. 2, Pp. 157-160, 8 July 1957.

2. Burcher, P.E. and Landrum, R.G. The Application of Dynamic Polarization to Earth's Field Nuclear Resonance Magnetometers, Master's Thesis, Naval Postgraduate School, Monterey, 1965.

3. Bowen, C.H. Jr., An Earth's Field Magnetometer That Utilizes the Free Precession of Protons, Master's Thesis, Naval Postgraduate School, Monterey, 1954.

4. RCA Linear Integrated Circuits, Technical Series IC - 41, Pp. 1-127, Radio Corporation of America, 1967.

5. GE Transistor Manual, 7th ed., Pp. 95-239, General Electric Company, 1964.

6. Lytel, A., Transistor Circuit Manual, 1st ed., Pp. 181-214, Howard W. Sams Photofact, 1961.

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Security Classification

DOCUMENT CONTROL DATA - R & D

(Security classification of title, body of abstract and indexing annotation must be entered when the overall report is classified)

1. ORIGINATING ACTIVITY (Corporate author)	2a. REPORT SECURITY CLASSIFICATION
Naval Postgraduate School Monterey, California	Unclassified
	2b. GROUP

3. REPORT TITLE

Marginal Oscillator for a Modified Free Precession Magnetometer

4. DESCRIPTIVE NOTES (Type of report and, inclusive dates)

Master's Thesis; June 1969

5. AUTHOR(S) (First name, middle initial, last name)

Jerome William Offenberg

6. REPORT DATE	7a. TOTAL NO. OF PAGES	7b. NO. OF REFS
June 1969	41	6

8a. CONTRACT OR GRANT NO.	9a. ORIGINATOR'S REPORT NUMBER(S)
b. PROJECT NO.	
c.	9b. OTHER REPORT NO(S) (Any other numbers that may be assigned this report)
d.	

10. DISTRIBUTION STATEMENT

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11. SUPPLEMENTARY NOTES	12. SPONSORING MILITARY ACTIVITY
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13. ABSTRACT

The primary interest of the Naval Postgraduate School concerns the use of a magnetometer for mine detection, anti-submarine warfare, salvaging and other related naval operations.

The original concept of a modified free precession magnetometer utilizing the Overhauser Effect was formulated by A. Abragam. Independent research on this magnetometer was conducted at the Naval Postgraduate School by P.E. Burcher and R.G. Landrum fulfilling requirements for their master's thesis in 1965. The objective of this thesis was to develop an improved marginal oscillator for the magnetometer.

DD FORM 1 NOV 65 1473 (PAGE 1)

S/N 0101-807-6811

Security Classification

A-31408

Security Classification

14 KEY WORDS	LINK A		LINK B		LINK C	
	ROLE	WT	ROLE	WT	ROLE	WT
Marginal Oscillator						
Positive Feedback Amplifier						
Modified - Free Precession Magnetometer						
Overhauser Effect Applied to Free Precession Magnetometer						

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S/N 0101-807-6821

Security Classification

A-31409